

Electromagnetic simulations for Magnetic Resonance radio frequency coils

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Abstract— Magnetic Resonance Imaging and Magnetic Resonance Spectroscopy are non-invasive diagnostic techniques employing magnetic fields without exposure to ionizing radiation.

Radiofrequency coils are key components in Magnetic Resonance systems since the use of coils which fit around parts of the body to be imaged is necessary for obtaining high-quality images and spectrums. The purpose of the transmitter coil is to produce a highly homogeneous alternating magnetic field in a wide field of view while the function of the receiver coil is to maximize signal detection while minimizing the noise. For optimizing coils performances for a given application, an accurate design process must be followed.

This work reviews methods for coil design and simulation, starting from coil characterization by using its equivalent RLC electric circuit and the use of magnetostatic approach for magnetic field distribution calculation, useful for the design of low frequency coils. For high-frequency tuned coils, the use of numerical methods is very attractive because complex structures such as part of human body can be incorporated in the computational space, permitting the study of the electromagnetic interaction between the coil and the biological phantom. The paper contains an overview of the state-of-art regarding numerical method applications.

Index Terms— Electromagnetic simulations, Magnetic Resonance, Radiofrequency coils

I. INTRODUCTION

Magnetic Resonance Imaging (MRI) and Magnetic Resonance Spectroscopy (MRS) are non-ionizing and non-invasive diagnostic techniques based on Nuclear Magnetic Resonance (NMR) phenomenon. MRI employs the signal from protons for producing anatomic images, thanks to ¹H species high natural concentration in the human body as part

of the water molecule, while MRS determines the concentration of different chem

ical species in the examined tissue. The basic physical principle is related to the behaviour of atomic nuclei with a magnetic moment, which, when placed in a static magnetic field (B_0) precess around its direction at a specific frequency, defined Larmor frequency:

$$f_0 = \frac{\gamma}{2\pi} B_0 \quad (1)$$

where γ is the gyromagnetic ratio, characteristic of a given nucleus [1].

Each magnetic moment will have a component along the static field direction and a component in the orthogonal plane. All the nuclei within a region characterized by the same static field intensity will precess at the same frequency, but out of phase. In this condition, the net magnetization in the perpendicular plane will be zero, being random the direction along this plane of each magnetic moment. For studying the tissue characteristics, the nuclei in the static field are irradiated by a radiofrequency (RF) magnetic field (B_1) whose direction is perpendicular to B_0 field direction and whose frequency is close to the Larmor frequency. B_1 (more in detail, its circular polarized component rotating as precession, i.e. B_1^+) will cause the nuclear magnetization of atoms to make an excursion from the static field and to become in-phase with each other, creating a net magnetization to a desired angle away from its equilibrium position. Applying an RF pulse for a duration τ seconds results in an angular deflection (flip-angle) [2]:

$$\theta = \gamma \int_0^{\tau} B_1(t) dt \quad (2)$$

It is frequently important to achieve the maximum signal received by the MR system ($\theta=90^\circ$) and this is accomplished by the so-called “90° pulse”, whose duration in ms provides B_1 field intensity in the range of μT .

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When the B_1 is switched off, the nuclei will return to equilibrium state while emitting signals at the same B_1 frequency. Tissue characteristics can be achieved by studying the atoms magnetic moment dynamics, since the way each nucleus returns to equilibrium state is related to the kind and position of nearby atoms, to the surrounding environment and to other externally applied magnetic fields (gradient field) employed for encoding nuclei position in space. Table 1 shows gyromagnetic ratio values for typical nuclei employed in MRI/MRS [1]. B_0 field strengths from 0.5 to 7 T correspond to B_1 field frequency in the range of 21-298 MHz for ^1H MR experiments.

Nucleus	^1H	^{19}F	^{31}P	^{23}Na	^{13}C
$\gamma/2\pi$ (MHz/T)	42.58	40.08	17.25	11.26	10.71

Tab. 1 Gyromagnetic ratio values for typical nuclei employed in MRI/MRS

In MR systems, the RF field is generated and picked-up by transmitter and receiver coils, respectively [1]. The purpose of the transmitter coil is to produce a highly homogeneous field in a wide field of view (FOV), since the extension of the region under analysis is not known a priori. Therefore, transmit coils are usually large to optimize the field homogeneity and include a significant volume of tissue. The function of the receiver coil is to maximize signal detection while minimizing the noise; for this purpose, coil dimensions have to be minimized. Finally, both transmit and receive coils must be adapted to the specific goal and to the sample dimensions. Nonetheless, they have to keep good performances with slightly different sample geometries, e.g. with different subjects.

This review describes radiofrequency coils theory and classification, and the different methods for coil performance estimation. A brief description of coil performance workbench measurement and a summary of future developments in coils technology and simulation is included.

II. RADIOFREQUENCY COILS CLASSIFICATION

RF coils can be categorized into two groups, according to their shapes [3]. The first group contains the volume coils, constituted mainly by Helmholtz and birdcage coils, that are often used both for transmission and reception since they can generate an uniform field in a large region surrounding the sample. The Helmholtz coil is constituted by a couple of two identical circular loops separated by a distance equal to the loop radius (see Fig. 1 left) and it is able to generate an uniform field in the vicinity of their midpoint [4].

The birdcage coil, popular due to its possibility to generate a highly homogenous RF magnetic field with a high signal-to-noise ratio (SNR), is constituted by N legs connected at each end to two circular loops (end rings) (see Fig. 1 right) [5]. The

birdcage coil is characterized by more than one resonant mode, since an N legs coil will supply $N/2+1$ distinct resonant modes [6], although the useful resonant mode is the one that guarantees the best magnetic field homogeneity.

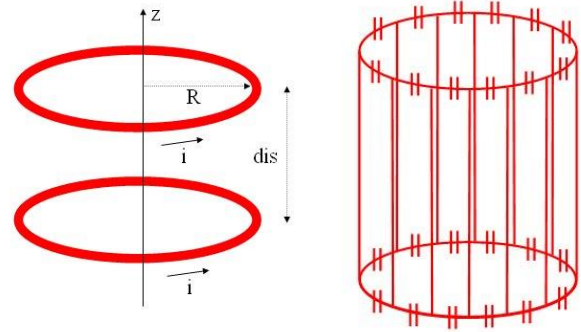


Fig. 1 Helmholtz coil (left) and birdcage coil (right)

The second group are the surface coils, which includes loop coils of various shapes (see Fig. 2 left). These coils are much smaller than the volume coils and provide higher SNR, but due to the relatively poor field homogeneity they are mainly used as receive coils.

In a MR experiment, the ideal coil setup should comprise the use of two different coils: a transmit coil, in order to achieve an efficient excitation in a large volume, and a receive coil, for maximizing the local sensitivity. However, surface coils usable FOV is limited to the sensitive region size, whereas it is desirable to have a large FOV because the region of interest is often not known prior to the first scan.

Phased-array coils [7] permits to provide a large region of sensitivity, similarly to that obtained with volume coils, and a high SNR, usually associated with surface coils. Each array element, generally constituted by circular or rectangular loops, is connected to an independent receiver channel and the outputs from these channels are combined in an optimal manner with a phase correction dependent on the point in space from which the signal is originated (see Fig. 2 right).

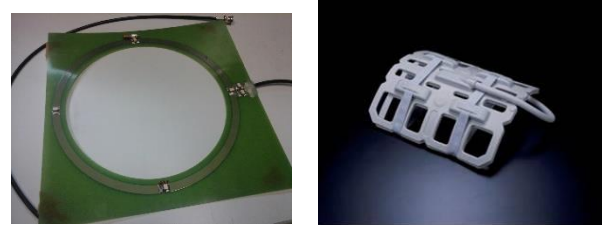


Fig. 2 Circular surface coil (left) and phased-array coil (right)

Mutual coupling between coil elements can be achieved by overlapping the adjacent coils to yield zero mutual inductance while for reducing the interference between the overlapped couple and the other coils in the array, each coil has to be connected to a low input impedance preamplifier [8].

Recently, the implementation of parallel imaging techniques in combination with dedicated phased-array coils significantly shortened data acquisition [9]. Although multi-channel coils are commonly used in receive mode with a body RF transmit coil, recently transmitter phased array coil systems were proposed [10], with the advantage to lead to a highly uniform tissue excitation. More recently, the need to minimize losses in the cables and crosstalks between multi-channel coils lead to develop digital RF coils, where the signal is sampled and digitalized directly in the coil chassis, by permitting SNR improvement [11]. Such digital RF coils are today available on some “top-level” scanners from several vendors.

For applications in vertical B_0 MRI systems (e.g. open magnets), coil geometry has to be designed such that the currents flowing in it produce a B_1 field that is substantially a transverse RF field. These transverse field RF coil are constituted by linear elements parallel to each other and connected by return current paths (butterfly and figure-of-eight coils), both in single or array configurations [12].

III. RADIOFREQUENCY COIL AS RLC CIRCUIT

RF coil can be schematized by an equivalent RLC circuit whose current which flows on it is maximized at the Larmor frequency (Fig. 3). As according to the reciprocity theorem [13], V can be the voltage source (transmit coil) or the sample-induced voltage (receive coil). L is the system inductance which takes into account for the energy that can be stored in the magnetic field and it is related to the conductors size and geometry. C is the system capacitance and is mainly resulting from the contribution of discrete capacitors.

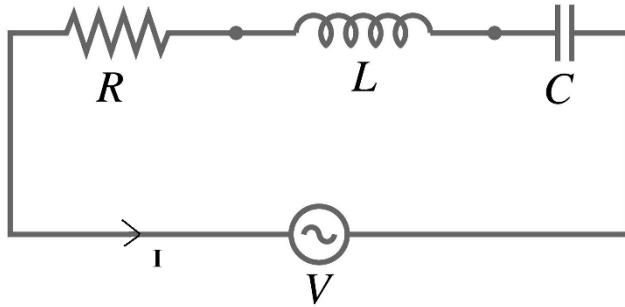


Fig. 3 RLC equivalent circuit of a radiofrequency coil

Energy exchange between magnetic and electric field might alternate in time with maximum efficiency at the resonant frequency. By applying Kirchhoff law, the circuit resonant frequency can be calculated as:

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \quad (3)$$

The resistance R is the sum of all the resistances that can be associated to loss mechanism within the conductors and within the sample [14]. In particular

$$R_{tot} = R_{coil} + R_{sample} + R_{extra} \quad (4)$$

R_{coil} takes into account for the losses within the coil conductors and depends on the conductor geometry. R_{sample} are the sample losses caused by RF currents, induced by the fluctuating magnetic field, and by electric fields in the sample, mainly generated by the coil capacitors. R_{extra} include radiative and tuning capacitors losses, which can be neglected in many applications.

The definition of the coil quality factor provides a quantitative measure of circuit quality, as [1]:

$$Q = 2\pi \frac{\text{energy stored in } i\text{-th cycle}}{\text{energy dissipated during } i\text{-th cycle}} \quad (5)$$

which can be expressed in terms of circuit parameters as:

$$Q = \frac{2\pi f_0 L}{R_{tot}} = \frac{1}{R_{tot}} \sqrt{\frac{L}{C}} \quad (6)$$

In dependence on their cross-sectional shape, conductors used for coils building can be categorized into two groups: wires (cylindrical rod shapes) and strips (rectangular shapes). While conducting wires size is defined by their radius, conducting strips are characterized by width and thickness (see Fig. 4).

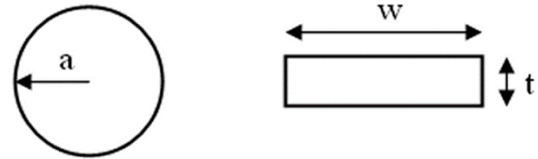


Fig. 4 Wire (left) and strip (right) conductors

IV. COIL RESISTANCE

An alternating current (AC) flowing in a conductor is not uniformly distributed across its section but it is confined in a region near the surface whose thickness (penetration depth) δ is given by [15]:

$$\delta = \sqrt{\frac{\rho}{\pi f \mu_0}} \quad (7)$$

where ρ is the conductor resistivity ($\rho = 1.68 \cdot 10^{-8} \text{ m} \cdot \Omega$ for copper), f is the coil tuning frequency and μ_0 is the free space permeability ($4\pi \cdot 10^{-7} \text{ Henry per meter}$). The conductor volume crossed by the RF current is limited by penetration depth value given the so called the “classical skin effect” [16].

The coil resistance can be estimated by using the classic formula $R = \rho l / S$ which takes into account for the conducting pathway geometry, where l and S are the total conductor length and cross-sectional area respectively [17].

Coil resistance for wire conductors

For a wire conductor, if the wire radius a is much greater than δ , the conductor losses for unit length can be calculated as follows:

$$R_{\text{clas-wire}}(f) = \frac{\rho}{2\pi a \delta} \quad (8)$$

where the dependency on frequency is due to the penetration depth, as in Eq. (5).

Coil resistance for strip conductors

For strip conductors of w width, when the strip thickness t is greater than two times the penetration depth size, the conductor resistance can be evaluated as:

$$R_{\text{clas-strip}}(f) = \frac{\rho}{2w\delta} \quad (9)$$

otherwise, the current flows in the total conductor cross-sectional area and the resistance value has to be calculated as:

$$R_{\text{clas-strip}}(f) = \frac{\rho}{tw\delta} \quad (10)$$

In real cases the tendency of the current density to concentrate towards the surface is more marked at the points where the curvature is greatest. For instance, in an elliptic cross-section conductor, the current density is enhanced at the ends of the major axes with respect to the ends of the minor axes. Now, assuming the strip conductor as the limiting case of a very thin elliptic cylinder, the current concentration is expected to be enhanced at conductor edges. This phenomenon is called the “lateral skin effect” [18]. The coil resistance then should be considered as the sum of the lateral skin effect and the classical skin effect resistances [19]:

$$R_{\text{coil-strip}} = R_{\text{clas-strip}} + R_{\text{lat-strip}} \quad (11)$$

Obviously, the wire conductor resistance is equal to the classical skin effect resistance due to the absence of conductor edges in such conductor:

$$R_{\text{coil-wire}} = R_{\text{clas-wire}} \quad (12)$$

In a previous work [20] authors proposed a comparison between different birdcage coils, tuned at 7.66 MHz (^1H frequency at 0.18T), two of them constituted by strip conductors with different thickness and the third built using a wire conductor. The paper showed that the use of strips with the thickness much higher than the penetration depth allowed to increase the overall coil performance, although the best performance was provided by the coil made of wire.

In [19] authors propose a theoretical-experimental hybrid method, which permits to distinguish and to quantify the different skin effect contributions to the conductor resistance when the coil is built using strip conductor. For achieving this, two circular coil prototypes with identical inductance value were designed and built with identical dimensions (the first one constituted by a strip and the second one by wire conductors).

From quality factor measurements, using Eq. (6), the total loss resistances for both coils were estimated as:

$$R_{\text{tot-strip}} = \frac{2\pi f L}{Q_{\text{strip}}} \quad (13)$$

$$R_{\text{tot-wire}} = \frac{2\pi f L}{Q_{\text{wire}}} \quad (14)$$

As stated previously, the coil constituted by wire conductors is affected by only classical skin effect, therefore the term R_{extra} can be estimates as:

$$R_{\text{extra}} = R_{\text{tot-wire}} - R_{\text{clas-wire}} \quad (15)$$

while the strip coil resistance can be evaluated as:

$$R_{\text{coil-strip}} = R_{\text{tot-strip}} - R_{\text{extra}} \quad (16)$$

being the term R_{extra} the same for both coils, which are identical in sizes and tuned with the same capacitors.

By using Eq. (11), the contribution of the lateral skin effect can be evaluated as:

$$R_{\text{lat-strip}} = R_{\text{coil-strip}} - R_{\text{clas-strip}} \quad (17)$$

The test results of both coils, tuned at low frequency (5.6 MHz), showed that the wire coil provided better performance, with an increase of 59% in the Q factor with respect to the strip coil. Moreover, the two contributions of the different skin effects for the strip coil assumed very similar values, the classical skin effect and the lateral skin effect resistance contributions being equal to 51% and 49% of the strip coil resistance, respectively.

A more recent work [21] evaluated the coil resistance at different tuning frequencies usually used in clinical scanner (21–128 MHz, corresponding to 0.5–3 T static field).

The frequency dependent of the lateral skin effect resistance showed a proportionality to the square of the frequency and underlined that in all frequency range routinely used in MR clinical scanner the lateral skin effect is the dominant mechanism and can not be neglected especially in high frequencies. With the described method, the total coil resistance for a strip coil can be evaluated by using the classical formulation (Eqs. (9) and (10)), which takes into account for the classical skin effect, and successively, it is possible to “correct” the resistance value by taking into account for the lateral skin effect contribution. The ratio of $R_{\text{lat}}/R_{\text{clas}}$ has been computed as a function of the coil tuning frequency [21].

V. SAMPLE-INDUCED RESISTANCE

For estimating losses from sample-induced resistance, literature cited works used homogeneous infinitely long cylinders [22], spheres [23] or half-spaces [24] as approximations of the sample geometry.

In [25] authors proposed two different procedures for sample-induced resistance calculation in a real MR experiment.

Sample induced resistance estimation using vector potential calculation

The first method, which employs vector potential calculation (VPC) for power losses estimation, is based on the magnetostatic approach, and can be easily mathematically implemented for sample-induced resistance calculation of simple coils and sample geometries.

In dependence on the coil geometry and shape, sample-induced resistance estimation can be performed by using [26]:

$$R_{sample} = \sigma \omega^2 \int_{vol} A^* A dV \quad (18)$$

where σ is the sample conductivity, A is the magnetic vector potential produced by the current flowing into the coil and the integral is over the sample volume.

Literature describes various algorithms for vector potential calculation to be used in Eq. (18), assessed with Legendre polynomials [27], spherical harmonics [26], surface integrals [28] and Lipschitz-Hankel integral [29]. However, all these methods require tedious calculations while a simple method based on the quasi-static approach, which can be easily mathematically implemented for sample induced resistance calculation of simple coils such as the circular loop, can provide good accuracy.

The magnetic vector potential produced by a current flowing in circular loop of radius b can be calculated as [1]:

$$A(r) = \frac{\mu_0 I}{4\pi} \int_C \frac{dl}{R} \quad (19)$$

where $\mu_0 = 4\pi \cdot 10^{-7}$ henrys per meter (H/m) is the free space permeability, I is the electric current flowing in the C contour, dl is the infinitesimal vector tangential to C and R is the distance between the observation point with (x, y, z) coordinates and the conductor path (see Fig. 5).

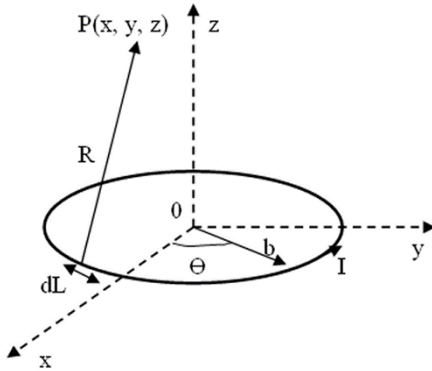


Fig. 5 Circular loop

The infinitesimal segment components in the conductor path are:

$$dl \equiv (dx, dy, dz) \equiv b(-\sin \theta, \cos \theta, 0) d\theta \quad (20)$$

and the coordinates of R vector are:

$$R \equiv (x - b \cos \theta, y - b \sin \theta, z) \quad (21)$$

Using Eq. (19) the three components of the magnetic vector potential can be calculated as:

$$A_x = \frac{\mu_0 I}{4\pi} \int_0^{2\pi} \frac{-b \sin \theta d\theta}{[(x - b \cos \theta)^2 + (y - b \sin \theta)^2 + z^2]^{\frac{1}{2}}} \quad (22)$$

$$A_y = \frac{\mu_0 I}{4\pi} \int_0^{2\pi} \frac{b \cos \theta d\theta}{[(x - b \cos \theta)^2 + (y - b \sin \theta)^2 + z^2]^{\frac{1}{2}}} \quad (23)$$

$$A_z = 0 \quad (24)$$

By integrating in a volume identical to the sample dimensions (FOV_x, FOV_y and FOV_z, with FOV=Field Of View), the sample-induced resistance calculation can be performed as:

$$R_{sample} = \sigma \omega^2 \int_0^{FOV_x} \int_0^{FOV_y} \int_0^{FOV_z} (A_x^2 + A_y^2 + A_z^2) dx dy dz \quad (25)$$

This method can be easily mathematically implemented in a code and is able to provide the results in short time. However, the integral in Eq. (25) can be easily performed only when the coil has a very simple geometry (e.g. circular, square or solenoidal) and the sample can be approximated by a simplified model, while in real MR experiments, both shape and dimensions can be very different respect to these coil and sample models.

The accuracy of this approach was performed with workbench tests on different coil sizes and tuning frequencies [25]. Since the vector potential is estimated with magnetostatic approach, its use would result in a significant error when the coil and sample dimensions are not small compared to the wavelength, i.e. at high field MR imaging.

Therefore, an accurate simulation can be performed only using a numerical method.

Sample induced resistance estimation using FDTD algorithm

This approach uses a numerical method which implements Finite-Difference Time-Domain (FDTD) algorithm [30]. Since it permits to incorporate in the computational space complex structures such as part of human body, the simulation of systems with various geometries, without by using approximations in sample and coil geometries, are allowed [31].

The algorithm employs resonant circuits theory [1], starting from the circuit quality factor defined in Eq. (6). In the numerical simulation, the resonant circuit was perturbed with a Gaussian pulse and the damping of the voltage oscillation was observed. Successively, the system Q factor can be evaluated from the energy stored by a capacitor, which is proportional to the square of the voltage across its layers, as:

$$Q = 2\pi \frac{V_i^2}{V_i^2 - V_{i+1}^2} \quad (26)$$

where V_i and V_{i+1} are, respectively, the voltage at i -cycle and at $(i+1)$ -cycle.

Having performed the simulation by considering the coil conductor constituted by Perfect Electric Conductor (PEC), the coil resistance vanishes completely and the series resistance value, which takes into account the energy dissipated in the system, can be calculated as:

$$R_s = \frac{2\pi f_0 L}{Q} \quad (27)$$

where f_0 is the coil resonant frequency and L is the coil inductance value.

Although the numerical simulation would require long computational time especially for low frequency tuned coils, accurate models of sample-coil interaction can be obtained at all frequencies routinely used in MRI applications for different coil and sample geometries [32, 33, 34].

More recently, a description of how the position of the sample inside the coil influences the sample-induced resistance value and an innovative method based on the exponential fitting on voltage oscillation damping has been described [35].

VI. COIL INDUCTANCE

The self-inductance of a conductor, which is fed with a current density J is evaluated by means of the following expression [1]:

$$L = \frac{\mu_0}{4\pi I^2} \iiint_V \iiint_V \frac{J(r) \cdot J(r')}{R} dv dv' \quad (28)$$

where μ_0 is the permeability of free space, I represents the total current in the conductor, V is the conductor volume and $R = |r - r'|$. This formulation is useful for the design of coils constituted by linear and circular conductor segments [36].

Straight conductors

The application of Eq. (28) to the calculation of the inductance of a strip of width w , length l and negligible thickness provides:

$$L = \frac{\mu_0}{4\pi w^2} \int_{-w/2}^{w/2} \int_{-l/2}^{l/2} \int_{-w/2}^{w/2} \int_{-l/2}^{l/2} \frac{1}{R} dx dy dx' dy' \quad (29)$$

with $R = \sqrt{(x - x')^2 + (y - y')^2}$ (see Fig. 6).

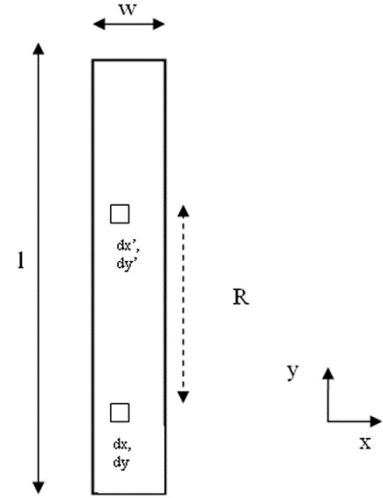


Fig. 6 Strip straight conductor

However, by assuming the current uniformly distributed over its surface, the evaluation of the inductance can be performed with a simple formula [1]:

$$L = \frac{\mu_0 l}{2\pi} \left(\ln \frac{2l}{w} + \frac{1}{2} \right) \quad (30)$$

whose accuracy increases as l is large with respect to w .

In the case of wire conductor of radius a and length l , by assuming that the current is uniformly distributed over its surface, the inductance calculation can be approximated by the following formula [1]:

$$L = \frac{\mu_0 l}{2\pi} \left(\ln \frac{2l}{a} - 1 \right) \quad (31)$$

whose accuracy increases as l increases with respect to a .

Now it is possible to find a simple relationship between strip and wire conductors sizes that permits to maintain the same inductance value.

These can be achieved by equalling the Eq. (30) and (31) and assuming the same length l for the two conductors [19]:

$$\frac{\mu_0 l}{2\pi} \left(\ln \frac{2l}{w} + \frac{1}{2} \right) = \frac{\mu_0 l}{2\pi} \left(\ln \frac{2l}{a} - 1 \right) \quad (32)$$

Carrying out simple mathematical operations, the relationship that allows to evaluate the equivalent width w of a wire of radius a , is:

$$w = 4.482 a \quad (33)$$

Circular segment conductors

The value of the inductance for a circular segment of radius b constituted by a w width strip can be calculated as outlined in Fig. 7.

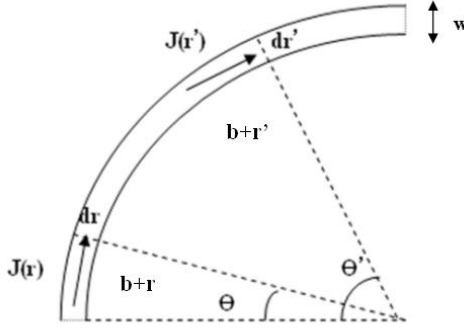


Fig. 7 Circular segment strip conductor

The infinitesimal surfaces are given by the following expressions [36]:

$$ds = \frac{\pi(b+r+dr)^2 - \pi(b+r)^2}{2\pi} d\theta \equiv (b+r) dr d\theta$$

$$ds' = \frac{\pi(b+r'+dr')^2 - \pi(b+r')^2}{2\pi} d\theta' \equiv (b+r') dr' d\theta' \quad (34)$$

with the points coordinates $P(r)$ and $P(r')$ that are given by:

$$P(r) \equiv (x, y) \equiv [(b+r)\cos\theta, (b+r)\sin\theta]$$

$$P(r') \equiv (x', y') \equiv [(b+r')\cos\theta', (b+r')\sin\theta'] \quad (35)$$

The term in (28) given by the scalar product of the current densities divided by the square of the total current becomes:

$$\frac{J(r) \cdot J(r')}{I^2} = \frac{1}{w^2} \cos(\theta' - \theta) \quad (36)$$

Therefore, the distance R between the two points is given by:

$$R = \sqrt{(x-x')^2 + (y-y')^2} =$$

$$\sqrt{[(b+r)\cos\theta - (b+r')\cos\theta']^2 + [(b+r)\sin\theta - (b+r')\sin\theta']^2} \quad (37)$$

The self-inductance of a circular loop can be evaluated by integrating dr and dr' from 0 to w (strip width of the conductor) and $d\theta$ and $d\theta'$ are integrated from 0 to 2π :

$$L = \frac{\mu_0}{4\pi w^2} \int_{-w/2}^{w/2} \int_0^{2\pi} \int_{-w/2}^{w/2} \int_0^{2\pi} \frac{\cos(\theta' - \theta)(b+r)(b+r')}{R} dr dr' d\theta d\theta' \quad (38)$$

The same theory can be employed for the inductance calculation of circular segments constituted by a radius wire conductors (see Fig. 8: (left) section and (right) upper view (b indicates the segment radius)).

Remembering Eq. (28), the surface infinitesimal in spherical coordinates, are given by [6]:

$$ds = a(b+a\cos\phi) d\phi d\theta$$

$$ds' = a(b+a\cos\phi') d\phi' d\theta' \quad (39)$$

while the coordinates of the points k and k' are:

$$k \equiv (x, y, z)$$

$$\equiv (b\cos\theta + a\cos\phi\cos\theta, b\sin\theta + a\cos\phi\sin\theta, a\sin\phi)$$

$$k' \equiv (x', y', z')$$

$$\equiv (b\cos\theta' + a\cos\phi'\cos\theta', b\sin\theta' + a\cos\phi'\sin\theta', a\sin\phi') \quad (40)$$

The term given by the scalar product of the current density divided by the square of the total current becomes:

$$\frac{J(r) \cdot J(r')}{I^2} = \frac{1}{(2\pi a)^2} \cos(\theta' - \theta) \quad (41)$$

The distance R between the two points is given by:

$$R = \sqrt{(x-x')^2 + (y-y')^2 + (z-z')^2} \quad (42)$$

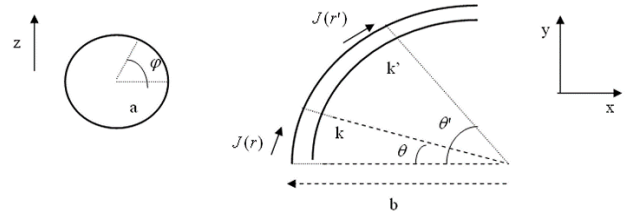


Fig. 8 Circular segment wire conductor

The final expression that permits the calculation of the inductance value of a circular loop constituted by wire conductor is made of a quadruple integral, in which $d\phi$ and $d\phi'$ are integrated from 0 to 2π (on the section of the conductor that constitutes the segment), while $d\theta$ and $d\theta'$ are integrated from 0 to 2π :

$$L = \frac{\mu_0}{4\pi(2\pi a)^2} \int_0^{2\pi} \int_0^{2\pi} \int_0^{2\pi} \int_0^{2\pi} \frac{\cos(\theta' - \theta)}{b} a^2 (b+a\cos\phi) (b+a\cos\phi') d\phi' d\theta' d\phi d\theta \quad (43)$$

However, the case of the conductors with wire geometry can be even approximately studied with the same theory used for the calculation of the inductances of the strip conductors and by using Eq. (33).

Mutual inductance

A similar expression to Eq. (28) can be used for mutual inductance evaluation between conductors. If I_1 and I_2 are the total currents in the V_1 and V_2 conductor volumes with current density J_1 and J_2 , the expression of the mutual inductance between the two conductors is given by [1]:

$$M_{12} = M_{21} = \frac{\mu_0}{4\pi I_1 I_2} \iiint_{V_1} \iiint_{V_2} \frac{J_1(r) \cdot J_2(r')}{R} dv dv' \quad (44)$$

VII. COIL TUNING CAPACITANCE

The best design for capacitors used for coil tuning requires a dielectric with extremely low loss at the frequency of use. A

capacitor may be depicted by an equivalent circuit constituted by an ideal capacitance and an equivalent resistance placed in series, which takes into account for the total ohmic losses.

Yeung et al. [37] firstly described the importance of capacitors quality factor in coil design, by proposing the modification of the conductor legs of a birdcage employing integrated capacitors using very low-loss materials as dielectrics, without any effects on the coil magnetic field homogeneity.

In [20] authors evaluated the coil performances in dependence on tuning capacitors quality factor. Two birdcage coils, both tuned at 7.66 MHz, were designed for application in a low magnetic field imaging scanner (0.18T). In particular, Q factor, ratio between unloaded and loaded Q and coil sensitivity were measured with workbench for the two coil, the first tuned by using commercially available ceramics capacitors (low quality factor) and the second with nonmagnetic ceramic chip capacitors (ATC 100B - American Technical Ceramics, USA). These last capacitors have high Q (> 10.000 at 1 MHz) and a sufficiently high breakdown voltage. The test showed that the coil which employed standard-quality capacitors experienced very low performance while significant difference were performed by using high quality capacitors, since the Q factor increases of about ten times, the r factor doubles and the coil sensitivity increases of about three times. Really, the high Q factor obtained at very low frequency is mainly related to the the quality factor of the ATC capacitors, which is higher respect to those provided working at high frequency [38].

It is important to underline that for coil construction, the capacitance tolerance should be small, since deviations in a single capacitor can perturb the B_1 symmetry, especially for birdcage coil [39].

VIII. MAGNETOSTATIC THEORY

Magnetostatic theory implies a nearly static field assumption that holds only when the coil dimensions are much lower than the wavelength. It has been demonstrated useful for the design of low frequency coils constituted by linear and circular conductor segments [6, 36], where the calculation of the magnetic field generated by the currents along the coil conductors can be performed by subdividing the coil path in segments, in order to separately study their total magnetic field contribution.

The static magnetic field produced by a steady electric current I flowing in an arbitrary closed contour C can be calculated by using Biot-Savart law [1]:

$$B(r) = \frac{\mu_0 I}{4\pi} \int_C \frac{dL \wedge R}{R^3} \quad (45)$$

where $\mu_0 = 4\pi \cdot 10^{-7}$ Henry per meter (H/m) is the free space permeability, dL is the infinitesimal vector tangential to C and R is the distance between the conductor path and the observation point (see Fig. 9).

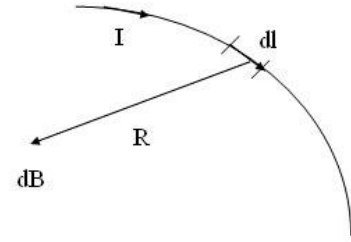


Fig. 9 Schematic representation of Biot-Savart law

Biot-Savart equation can be used for the simulation of the three-dimensional B_1 field distributions of the coils. The conductor sizes can be neglected respect to the wavelength, and the magnetic field generated by the currents can be evaluated considering conductors as very thin wires. Fig. 10 shows an example of a birdcage coil B_1 field pattern calculated with the software simulator described in [6].

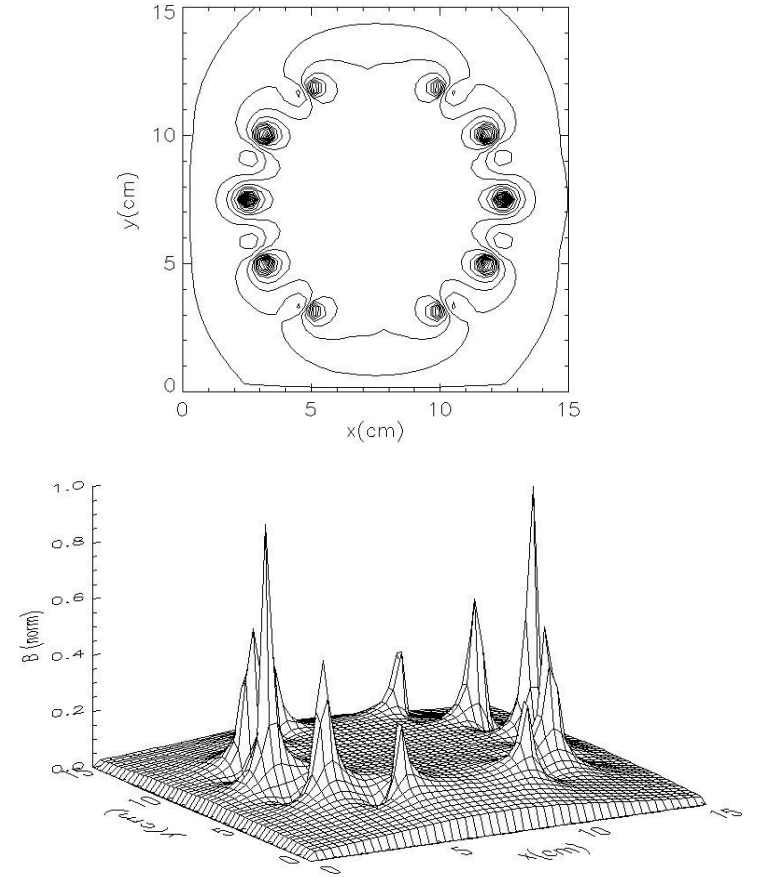


Fig. 10 Contour plot (above) and 3-dimensional representation (bottom) of a birdcage coil B_1 magnetic field pattern

IX. RF COILS SIMULATION METHODS

As frequency increases, RF fields interact more strongly with the sample, i.e. the human tissues, rendering quasi-static approaches no more reliable for electromagnetic characterization. Customized analytical methods can be

developed to investigate this electromagnetic characterization using simplified geometries and assumptions [40, 41]. For example, in [40] an approach based on the static solution for homogeneous cylinders illuminated by line current and homogeneous sphere illuminated by a loop is given, while [41] is based on the solution for of a-c current line (dynamic solution) and it is suited for multi-layered cylinders. However, the complex interactions between coils and sample cannot be usually solved using analytical methods; it follows that numerical methods based on solutions to Maxwell's equations are required [1]. In this context, many different numerical methods can be employed, including the Finite-Difference Time-Domain (FDTD) [42-46] the Finite Element Methods (FEM) [47, 48], the Method of Moments (MoM) [44, 49, 51]. In [44] a hybridization of MoM and FDTD is proposed, while more recently a number of works on hybridization between MoM/FEM and FDTD obtained growing attention [52, 54]. Finally, some authors have also successfully applied the Transmission Line Method (TLM) [55, 56] which derives also from differential Maxwell's equation in time-domain; thus, it can be considered a variation of FDTD.

FDTD, or its version involving the integral form of Maxwell's equations in time domain, i.e. the Finite Integration Technique (FIT), has been widely used for electromagnetic characterization of low and high field MR coil loaded with human-like phantom. However, when applied to volume coils having multi-mode resonances, it suffers of convergence problem. To overcome this difficulty many authors use a simplified geometry (without capacitor) which is enforced to operate at the RF and accordingly to the appropriate single-mode resonance [42, 57]. However, the aforementioned approach cannot take into account the effects of high-order modes, leading to an under-estimation of the Specific Absorption Rate (SAR) [58]. FDTD has also some critical issues when dealing with structures having small radius of curvature, i.e. the rods or tube legs of volume resonators.

FEM and MoM are both frequency-domain procedures; thus they are both inherently quicker than FDTD and FIT (time-domain procedures). However, MoM suffers in treating anatomic human models and it usually limits its applicability to homogeneous loads only. Concerning memory requirements, in FEM they scale proportional to the geometry, the frequency and the size of the surrounding space (i.e. for electrically small objects the space is very large). Instead, in MoM, memory requirements scale proportional to the geometry and the frequency, being the induced currents the unknowns of the problem (source method).

From this overview, it is possible to point out that each one of the mentioned numerical techniques permits the electromagnetic characterization MR coil, each one having its own vantages and drawbacks. In [59] a quantitative comparison of performances and a detailed evaluation of vantages and drawbacks has been performed for a 1.5T proton surface coil loaded with a cylindrical phantom, a 7T dual tuned (1H/31P) surface coil loaded with a cylindrical phantom, a 7T unloaded proton volume coil.

Concerning the 7T dual tuned surface coil shown in Fig. 11 and Fig. 12, we observed an excellent agreement in the prediction of the conditions of tuning and matching between

measurements and the three methods, as given in Table 2. More in details, the best agreement has been obtained through FIT. Concerning the magnetic end electric field maps, the average percentage difference between the methods is less than 1% in all cases, showing a substantially perfect agreement between the results of the MoM, FEM and FIT (see Fig. 12). It follows that the three methods are all able to provide accurate results, and all the methods are able to highlight the effect of the sample at high frequency. However, there are substantial differences in the duration of the simulation and in the amount of memory allocated. The MoM is faster and it requires less memory since the unknowns are the induced currents and the field calculation is done (and stored) just at the required frequency. MoM can be usually applied when a homogeneous load is considered. The FIT simulation requires much longer time; however, if an inhomogeneous humanoid sample is required as load, FIT method has to be used. We can therefore say that, for this geometries, the frequency domain methods, especially MoM, are suitable for simulations performed in the prototype stage, as they allow an immediate comparison being limited the duration of the simulation, while the FIT is more suitable for a simulation aimed at evaluating the SAR and magnetic field inside a complex humanoid sample (at the expense of an increased duration of the simulation).

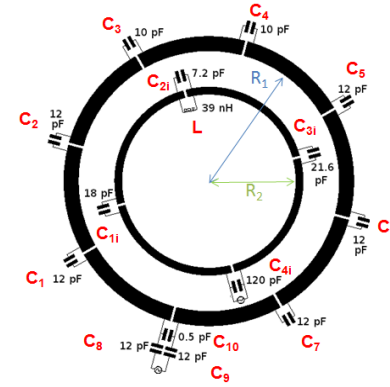


Fig. 11 7T dual tuned surface coil for 1H and 31P. The design of the dual tuned coil consists of two concentric loops, where $R_1 = 4.5$ cm and $R_2 = 3$ cm are the internal radius of the outer and the inner loop respectively. Each loop has been constructed with copper strips of width $w_1 = 0.5$ cm and $w_2 = 0.25$ cm respectively. The strip's thickness is equal to $35 \mu\text{m}$.

Quantity	Measure	MoM	FEM	FIT
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Frequency, 1H	298.6 MHz	6%	4%	2%
Matching, 1H	-47 dB	0.4%	1 %	10 %
Frequency, 31P	120.5	5 %	5 %	4 %
Matching, 31P	-24 dB	0.1 %	1 %	2 %
Coupling at H freq.	-65 dB	0.01 %	0.01 %	0.01 %
Coupling at P freq.	-13 dB	0.2 %	1%	0.6 %
Simulation time	/	15 m	2 h 57 m	17 h 6 m

Tab. 2 Comparison between measurements and the simulations (with the three different methods) of the dual tuned coil. The tuning and matching of both channels and decoupling between channels has been reported. Intel i7 920 with 12Gb of RAM memory has been used

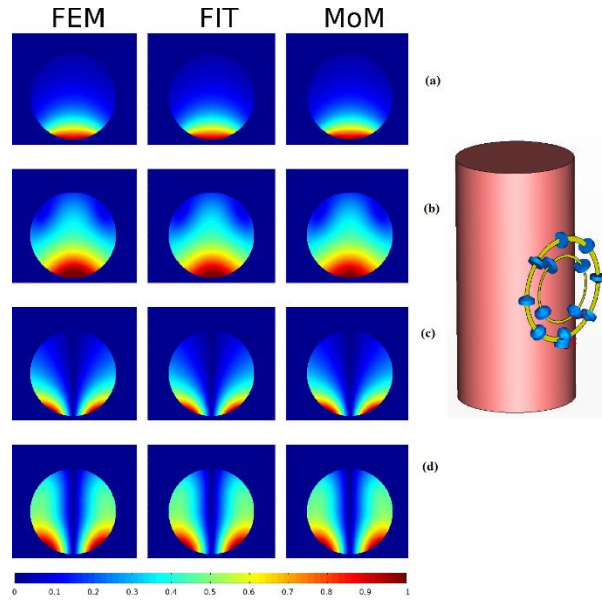


Fig. 12 (a) - (b) Transverse magnetic field simulated at 298 MHz (1H Larmor frequency) and at 120.6 MHz (31P Larmor frequency) respectively on a transverse plane of the phantom; (c) - (d) Electric field simulated at 298 MHz and at 120.6 MHz respectively on a transverse plane of the phantom (homogeneous cylinder having $\sigma = 0.686$ S/m, $\epsilon_r = 79$). Reprinted with permission

X. FIELDS AND SAR PRODUCED BY THE RF COILS

The management of SAR is a critical issue in MR, especially at ultra-high field (UHF) B_0 field strengths. In fact, at UHF ($B_0 \geq 4T$), the energy deposition due to the RF field increases and its distribution inside the subject becomes extremely inhomogeneous and subject-dependent [60, 61]. The increment of the RF energy deposition and of its spatial variability at UHF is due to the higher operational frequency of the UHF MR system, which is equal to 298 MHz at 7.0T.

Fig. 13 shows the B_{1+} and point SAR distribution in volume coil at 64 MHz and 298 MHz using current line sources approach to model a M=16 element 1H birdcage coil operating in quadrature. The elements of the coil are placed

equally spaced along a circle of radius of 29.5 cm; the radius of the shield is 37.5 cm (see [41] for more details). The load is a 6.5 cm radius cylinder mimicking the white matter [62], i.e. with dielectric constant of 52 and conductivity of 0.55 S/m. From Fig. 13, it is possible to note:

- The B_{1+} magnitude, which has been normalized, is much more homogenous at 64 MHz (Fig. 13 a), i.e. at 1.5T, rather than at 298 MHz (Fig. 13 b), i.e. at 7T: this can be explained recalling that at 298 MHz the wavelength in the load becomes comparable with respect to the dimension of the load itself. From Fig. 13b, it is possible to note a focusing effect in the center, which leads to the well-know phenomenon of central brightening in UHF MR images [40, 41, 63, 65]
- To allow SAR comparison, the SAR distribution at 64 MHz (Fig. 13 c), i.e. at 1.5T, and at 298 MHz (Fig. 13 d), have been obtained after scaling to the achieve a correspondent B_{1+} of 1 μT (averaged on the slice). After such scaling it is possible to note an increment of the RF energy deposition with the frequency. The same effect has been observed in [65] and motivate a more careful SAR monitoring at UHF.

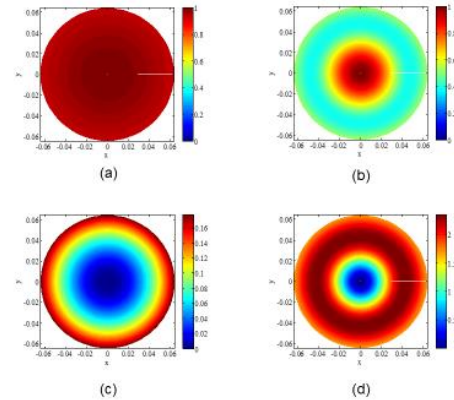


Fig. 13. B_{1+} magnitude (a, b) and point SAR (c, d) distribution in volume coil at 64 MHz (left column) and 298 MHz (right column) using line sources approach to model a M=16 element 1H birdcage coil operating in quadrature. The B_{1+} magnitudes have been normalized. SAR distributions [W/kg] have been obtained after scaling to the achieve a correspondent B_{1+} of 1 μT (averaged on the slice).

As stated previously, FEM and FIT are widely used for electromagnetic (EM) characterization of MR coil loaded with anatomic human models. In [63], a transmit-receive shielded 16 elements 1H high-pass birdcage head coil manufactured by Nova Medical (Wilmington, MA, USA), operating in quadrature at 298 MHz has been simulated using FIT in the CST MW Suite. A quadrature feed with 4 sources equally spaced azimuthally by $\pi/2$, with a relative electrical phase shift of $\pi/2$, was used. To validate the birdcage coil model, we performed a comparison between measured and simulated B_{1+} magnitude (see Fig. 14): more in detail, we measured the B_{1+} magnitude of a 0.1 M saline solution cylindrical phantom (radius=6 cm, height=24 cm), positioned in the coil as shown

in Fig. 14, by using the Bloch-Siegert Shift (BSS) method [64] on a GE MR950 7 T human system (GE HealthCare, Milwaukee, WI, USA) equipped with the birdcage coil described above. In the simulation, the 0.1 M saline solution load was modelled as a lossy homogeneous dielectric cylinder having permittivity of 80 and conductivity of 0.6 S/m. After birdcage model validation, we inserted as load a human head model derived from the $2 \times 2 \times 2$ mm³ voxel-size anatomic human model Ella (female, adult, 59 kg), Virtual population, ITIS foundation (see Fig. 15). B1+ was calculated in the axial slice crossing the corpus callosum and in the central saggittal slice, while the maximum local (10g) and global SAR [W/kg] were calculated in the entire head. In Fig. 15 we show the B1+ magnitude map and the local SAR on the saggittal slice; moreover, in the same figure, we report a table summarizing the maximum local (10g) and the global SAR obtained after scaling to the achieve a correspondent B1+ of 1 μ T (averaged on the two slice, separately). The maximum local and global SAR for a given sequence applied on a the aforementioned slices can then determined by a scaling factor which accounts for all the sequence-related parameters: TR, number and waveform of RF pulses, and the corresponding average nominal FAs (which are related to the average B1+).

For these simulations, approximately 12.5 million mesh nodes have been used (simulation time=15 h on one workstation using the Nvidia Quadro 6000 (Nvidia, Santa Clara, CA, USA) graphics processing unit (GPU) computing).

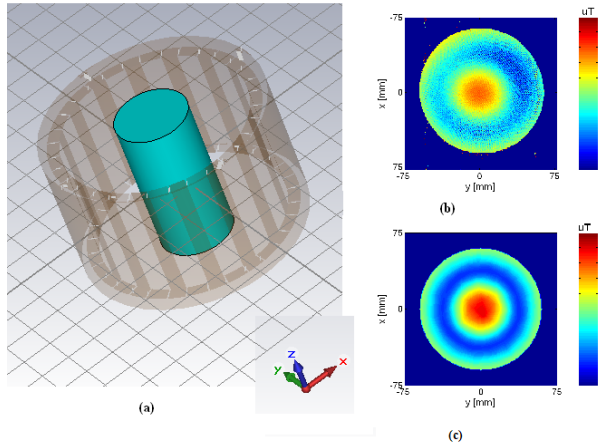


Fig. 14 a) Quadrature birdcage coil loaded with 0.1 M saline solution cylindrical phantom (radius of 6 cm, height of 24 cm); b) B1+ magnitude in the central z-plane, i.e., central axial slice, measured by using the BSS method and using a total input power of 0.1151 kW (note that this value represents the total input power at the birdcage section, since it has been obtained after taking into account all the cable losses in the RF chain); c) B1+ magnitude obtained through simulation. The central focusing effect is clearly visible in B1+ maps. B1+ is in μ T. Reprinted with permission.

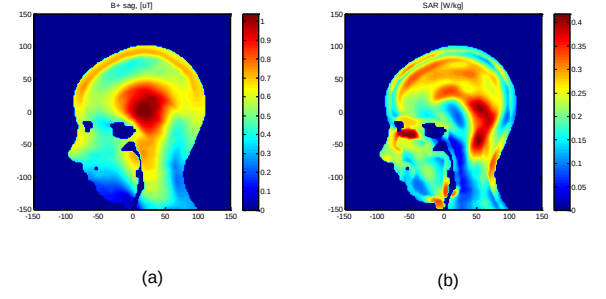


Fig. 15 B1+ magnitude map and the local SAR on the saggittal slice using a total input power of 4W. The central focusing effect (typical for the birdcage) is clearly visible in B1+ maps. B1+ is in μ T, SAR in W/kg. In the same figure, a table summarizing the maximum local (10g) and the global head SAR obtained after scaling to the achieve a correspondent B1+ of 1 μ T (averaged on the two slice, separately) is also given.

Each of the above-mentioned 3D full-wave numerical techniques permits the EM characterization of RF fields, namely the E and B1 field, and the calculation of the SAR distribution, as well as the S parameter matrix. In most cases, it is difficult to have access to the actual matching networks and to correctly reproduce them in the simulation; thus, it follows that the simulated S parameter matrix can be quite different from the actual one. RF fields and SAR distributions depend on the S parameter matrix: in [66] it has been shown that a simple phase-variation in some Sxy elements of the S matrix can lead to a SAR increase up to 30%. The use of 3D full-wave tools by itself becomes a severe limiting factor in the analysis multi-port MR coil when the anatomic human model is included in the simulation domain, because the full highly complex 3D EM problem must be solved for each matching and coupling condition (note that a multi-port MR coil can be either a volume coil driven in quadrature, or a coil for parallel transmission). Alternatively, two-way link between RF circuit and 3D EM simulation tools can be employed [67]: this enables simulation results from the RF circuit domain to be used to drive the 3D EM domain. Two-way link has been introduced in MR in [68].

XI. MEASUREMENT WITH WORKBENCH

The following definition provides a practical way for coil quality factor measurements (Fig. 11):

$$Q = \frac{f_o}{B_w} \quad (46)$$

where B_w is the -3 dB coil bandwidth.

A common parameter of coil performance evaluation is the ratio r between the quality factor of empty resonator (Q_{empty}) and resonator with the sample (Q_{sample}) [69]:

$$r = \frac{Q_{empty}}{Q_{sample}} = \frac{R_{coil} + R_{sample}}{R_{coil}} \quad (47)$$

r should be maximized to provide a maximal MRI signal-to-noise ratio, since $SNR \propto \sqrt{1 - \frac{1}{r}}$ [70].

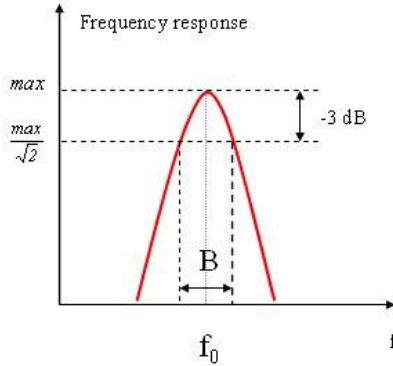


Fig. 16 Workbench coil quality factor estimation.

Resonant frequency and quality factor measurements can be easily performed by using a dual-loop probe, consisting of two pickup loops partially overlapped to minimise the mutual coupling between the elements, and a network analyser. Respect to other measurement setups which used direct contact with the coil, the dual-loop probe permits to place the transmitting and receiving probe elements approximately in the same place inside the under-test coil. The measurements are carried out by using one loop coil as a transmitter and the other one as a receiver. By using this set-up, the transmit loop is weakly coupled to the under-test coil, which in turn is weakly coupled to the receive loop, and the power transmitted to the receive loop is proportional to the amplitude of the oscillation in the coil under-test and, therefore, represents its frequency response [71].

Another important parameter which characterizes RF coil performance is the sensitivity, defined as the B_1 magnetic field induced by the RF coil at a given point per unit of supplied power P as follows [71]:

$$\eta = \frac{B_1}{\sqrt{P}} \quad (48)$$

The reciprocity theorem [13] allows the use of the same quantity defined in Eq. (7) to characterize both the transmit and the receive performance of an RF probe. It is important to note that maximizing the coil sensitivity will maximize also the SNR. Different magnetic field mapping methods that permit estimation of the coil efficiency at a fixed point in space are described in the literature. A good summary of these methods, classified between electromagnetic laboratory tests and MRI techniques, is available in a review [72] and in a more recent article dedicated to ^{13}C coils [73].

XII. OPEN ISSUES

The present overview is obviously not exhaustive: we simply wanted to highlight to readers, which are expert in EM, some important targets of radiofrequency research in MR. In this context, real time subject-specific local SAR evaluation represent one of the most demanding open issues. MR systems give an estimation of the global SAR in the subject under test during the exam; in fact, during MR exams, global SAR exposure has to be monitored so to remain below the regulatory limit imposed by the International Electrotechnical Commission (IEC), as given in regulation number IEC-60601-2-33; the limit for the head is 3.2 W/kg during any 6 minute time average. This measure is obtained by means of an empirical formulation, which takes into account the forward and reflected power during MR acquisition, and patient parameters [74]. The limit for the maximum head local SAR (10 W/kg during any 6 minute time average) is then supposed to be not overcome if the global SAR is below the limit.

However, the aforementioned measurement presents major drawbacks, including: i) the monitoring of forward and reflected power is performed in real time, but offers no capability for SAR prediction; ii) global SAR is determined by empirical formulation and thus it is not subject specific since subject anatomy and subject position with respect to the transmitting coil is not taken into account; iii) local SAR is not evaluated. Moreover, it has been shown that global SAR measurement routines differ from system to system: thus, they should not be taken as the primary and only measure to evaluate MR safety [75, 76].

SAR regulatory limits do not distinguish between adults (age ≥ 18 years) and children, but particular care must be paid when scanning juvenile subjects due to the lack of data on SAR exposure in children at UHF in the literature. While many B1+ field-mapping techniques exist (see [64] and correspondent references), subject-specific SAR measurements are not available in current MR systems; thus, electromagnetic simulations must be performed for RF fields and SAR analysis, as shown before. In Collins et al. [77] and van Lier et al. [78] authors have also addressed the relation between SAR and temperature using Pennes' bioheat equation.

SAR calculation relies on simulated data from human models, but such models cannot match precisely the anatomy of actual subjects. Subject-specific anatomic simulations models would require to previously acquire and segment images of each subject [79, 80]: however, such approach is rarely feasible and therefore models are used instead. This common practice introduces a mismatch between real and simulated data, which can be compensated by simulating and comparing different human models [63, 81]. Thus, it follows that real time subject-specific local SAR evaluation represent one of the key target of radiofrequency research in MR.

Fig. 14-16 clearly highlight the B1 field inhomogeneity at UHF; it follows that another open issues is related to the B1 field homogeneity enhancement. The phenomenon of central focusing effect in UHF MR images can be highlighted also by

looking at Fig. 17. Specifically, Fig. 17 shows the B1+ maps obtained on a GE MR950 7 T human system (GE HealthCare, Milwaukee, WI, USA) equipped with the birdcage coil described above, loading the coil with a 90 mm radius cylindrical phantom filled with oil (a) and saline water (b). Central focussing can be observed in Fig. 17 b, being the wavelength in the water comparable with respect to the dimension of the load itself.

Parallel transmission with multiple, independent RF coil elements has the most promising capability to influence the B1 distribution to achieve noticeable improvements in excitation and coverage in imaging [61, 82]. However, one major obstacle in parallel transmission is the question how to manage both global and local SAR, as multiple independent excitations will now superimpose inside the human body [61]. Another approach to increase B1 in regions of low RF transmit efficiency has been shown with high dielectric permittivity bags ($\epsilon_r = 100 \div 300$) placed between the RF coil and a region of low B1 [83], as shown in Fig. 18. However, more scientific effort is required to allow to find a strategy to determine the parameters (including size and position) of such bags which lead to optimal B1 increase, preferably without affecting local SAR.

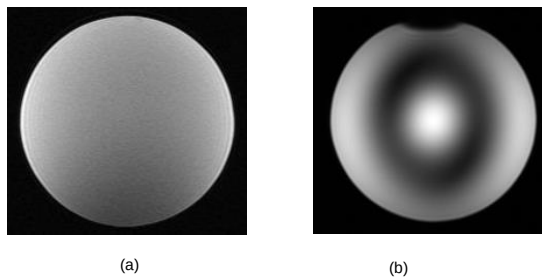


Fig. 17. B1+ maps (normalized) obtained on a GE MR950 7 T human system (GE HealthCare, Milwaukee, WI, USA) equipped with the birdcage coil described above, loading the coil with a 90 mm radius cylindrical phantom filled with oil (a) and saline water (b).

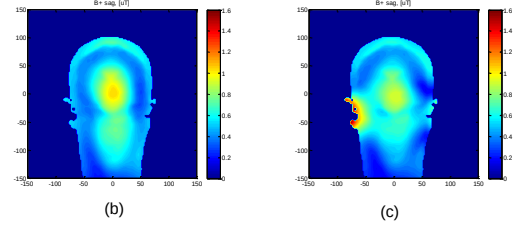
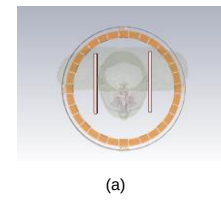


Fig. 18. Axial view of Ella inside the birdcage together with two rectangular dielectric BaTiO₄ bags, having $\epsilon_r=286$, $\sigma=0.44$; (b) Ella's coronal B1+ magnitude map (normalized) without bags; (c) Ella's coronal B1+ magnitude map (normalized) with the bags: B1+ enhancement in the ear region can be noticed.

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